

Mixed-Mode Third-Order Quadrature Oscillator Based on Single MCCFTA

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Abstract. This paper presents a new mixed-mode third-order quadrature oscillator based on new modified current-controlled current follower transconductance amplifier (MCCFTA). The proposed circuit employs one MCCFTA as active element and three grounded capacitors as passive components which is highly suitable for integrated circuit implementation. The condition and frequency of oscillations can be controlled orthogonally and electronically by adjusting the bias currents of the active device. The circuit provides four quadrature current outputs and two quadrature voltage outputs into one single topology, which can be classified as mixed-mode oscillator. In addition, four quadrature current output terminals possess high-impedance level which can be directly connected to the next stage without additional buffer circuits. The performance of the proposed structure has been verified through PSPICE simulators using 0.25 μm CMOS process from TSMC and experimental results are also investigated.

Keywords

Third-order quadrature oscillator, mixed-mode oscillator, modified current-controlled current follower transconductance amplifier (MCCFTA)

1. Introduction

At present, current-mode technique is very interesting approach due to the fact that it is easy as operating of arithmetic such as addition and subtraction signals, multiplication and division signals by a constant signal and potential to operate at lower supply voltage compared their voltage-mode circuits [1]. Regarding to a current-mode building block, the current differencing transconductance amplifier (CDTA) [2] is a really current-mode active building block, due to its input and output signals are current forms. The structure of this device is consisted of a unity-gain current source controlled by the difference of two inputs and an output transconductance amplifier providing electronic tuning capability through its transconductance gain (g_m). Therefore, CDTA is highly suitable for

realizing current-mode analog signal processing and CDTA-based circuits can also reduce number of passive resistors.

A quadrature oscillator (QO) usually provides sinusoids having a phase difference of 90° that is useful in communication and measurement systems such as quadrature mixers and single sideband generators for communication system [3], vector generators and selective voltmeters for measurements system [4]. Several QOs using CDTAs as active element have been proposed in the literature; see, for example [5–12]. However, some structures do not exploit the full capability of the CDTA where typically one of two input terminals of the CDTA is floated and not used [8–12]. Unfortunately, floating terminal may increase the area of chip when these CDTA-based QOs build as IC's forms and also may cause some noise injection into the monolithic circuit.

Recently, a new concept of active building block with one current input and two kinds of current outputs, the so-called “current follower transconductance amplifier (CFTA)”, has been introduced [13]. This device is modified from an original CDTA by removing the negative terminal. When subtraction current circuit as input stage is absent, the structure of CFTA is simple. Compared with CDTA, the number of transistor used for CFTA is lesser. Therefore, several CFTA-based analog signal processing circuits are reported [14–19]. The CFTA has been already used to realize QOs [20–26]. They, in conjunction, exhibit high potential for bring down the number of components. However, all structures are second-order QOs. This work focuses on the third-order QOs, which enjoys the requirement of QOs such as orthogonal adjustability and electronic tunability of the condition of oscillation (CO) and frequency of oscillation (FO), and uses grounded capacitors and minimum number of active elements. Because of high-order network the circuit provides better frequency response and quality, compared with lower-order network [27]. This mention has been expressed in [28] by mathematical formulation, especially, to confirm phase noise reduction. Compared with a second-order oscillator, the third-order oscillator (three-stage oscillator) also offers lower phase noise [29]. As a result, a number of third-order

QOs based on different active building blocks have been proposed [30–52]. The early system using operational transconductance amplifier (OTA) [30] enjoys electronic tuning capability, but its CO and FO are not decoupled and difficult to control. The QOs using active building blocks such as operational amplifier [31], second-generation current conveyor (CCII) [32–34], differential voltage current conveyor (DVCC) [35], operational transresistance amplifier (OTRA) [36], [37], [50], offer orthogonal control of CO and FO, but these structures lack the electronic tuning capability. A number of electronic-controlled third-order QOs have been reported using active building blocks such as current-controlled CCII (CCCII) [38], [39], current difference transconductance amplifier (CDTA) [40], [41], current-controlled CDTA (CCDTA) [42], current-controlled current conveyor transconductance amplifier (CCCTA) [43], [44], OTRA and MOS-C [45], DDCC and VDTA [46], [51], [52], differential voltage current conveyor transconductance amplifier (DVCCCTA) [47]. They exhibit high potential for enjoying up the electronic tunability. However, they require an excessive number of active components and the structures are not compact. The third-order oscillator QO based on log-domain technique is proposed in [48]. However, this structure is suitable only in bipolar technology. The structure in [49] proposed third-order QO using a single current-controlled current conveyor transconductance amplifier (CCCCTA), but the circuit does not exploit the full capability of the CCTA when y-terminal of CCTA is not used and attached to ground. Until now, there is no CCCFTA-third-order QO available in open literature.

Therefore, this paper presents a new mixed-mode third-order QO employing a modified current-controlled current follower transconductance amplifier (MCCFTA) and three grounded capacitors. The concept of MCCFTA is similar to conventional CFTA [13], except f -terminal provides parasitic resistance R_f that can be controlled by the bias current [53]. Identical z -copy terminal can be obtained using current-mirrors [13] and connecting transconductance amplifiers in parallel connection are available [54], [55]. Thus, MCCFTA is an active building block that provides the possibility for utilizing its resistance simulating element and its transconductance gains that can be electronically controlled through adjusting the bias currents. Therefore we have input parasitic resistance and transconductance gains (TAs) into a single MCCFTA for realizing a third-order QO with orthogonal control of the CO and FO. The proposed structure provides four quadrature current outputs and two quadrature voltage outputs. PSPICE simulation results are given to confirm the performance of the proposed structure. The comparison between the proposed third-order QO and previously third-order QOs is expressed as Tab. 1.

2. Proposed Circuit

The circuit symbol and the equivalent circuit of the MCCFTA are shown in Fig. 1 and CMOS implementation

of MCCFTA is shown Fig. 2. The ideal port relations of Fig. 1 can be expressed by

$$\begin{pmatrix} I_{z1} \\ I_{z2} \\ I_{zc} \\ V_f \\ I_{x1} \\ I_{x2} \end{pmatrix} = \begin{pmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 & 0 & 0 \\ R_f & 0 & 0 & 0 & 0 & 0 \\ 0 & \pm g_{m1} & 0 & 0 & 0 & 0 \\ 0 & 0 & \pm g_{m2} & 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} I_f \\ V_{z1} \\ V_{z2} \\ V_{zc} \\ V_{x1} \\ V_{x2} \end{pmatrix} \quad (1)$$

where R_f is the parasitic resistance at f -terminal and g_{m1} and g_{m2} are two TAs. The z_1 -, x_1 -terminals and g_{m1} are existing terminals and TA of conventional CCFTA [50]. The z_2 - and z_c -terminals are the z -copy terminals [13] of CCFTA that can be obtained by adding current mirrors in conventional CCFTA. By cascading g_{m2} in parallel connection, x_2 -terminal in CCFTA can be obtained and hence the name MCCFTA. It should be noted that connecting n TAs in parallel connection of MCCFTA is also possible. From the CMOS implementation of MCCFTA in Fig. 2, assuming that transistors M_1 to M_4 are matched and operated in saturation regions, the parasitic resistance at f -terminal (R_f) can be given as

$$R_f \cong \frac{1}{\sqrt{8\mu C_{ox} \left(\frac{W}{L}\right) I_{b1}}} \quad (2)$$

where μ is the carrier mobility, C_{ox} is the gate oxide capacitance per unit area, W and L are the channel width and length, respectively, of MOS transistor. From (2) and Fig. 2, the parasitic resistance R_f can be controlled by adjusting the bias current I_{b1} . This property makes it different from a conventional CFTA [13].

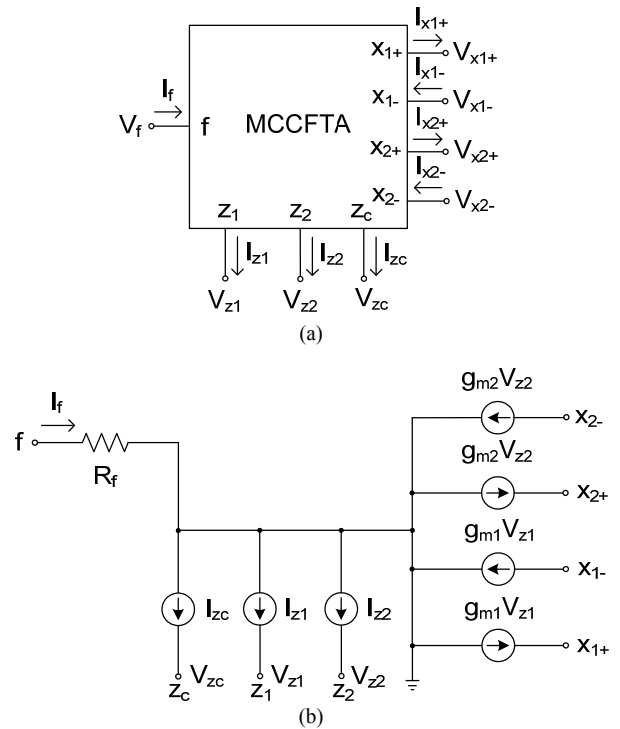


Fig. 1. MCCFTA: (a) electrical symbol, (b) equivalent circuit.

	Number of active elements	Number of resistor (R) & capacitor (C)	All-grounded passive element	Orthogonal control of CO and FO	Offer current and voltage outputs	Offer electronically tunable	Multi-phase generation	Type of dependence of FO on control/bias current/voltage	Availability of constant of output waveforms in dep. On tuning of FO	Range of output voltage/current (peak-to-peak) (Voltage) (Current)	Phased error of generate waveform (%) (Voltage) (Current)
Proposed QO	1-MCCFTA	3-C	Yes	Yes	Yes	Yes	Yes	Nonlinear	No	106-700 mV 20-160 μ A	0.3-1.89 0.5-5.78
Ref. [30] in 2002	3-OTA	3-C	Yes	Yes	No	Yes	Yes	Nonlinear	No	-	-
Ref. [31] in 2011	3-op-amp	3-C, 5-R	No	Yes	No	No	No	Nonlinear	No	-	-
Ref. [32] in 2005	3-CCIIIs	3-C, 5-R	Yes	Yes	No	No	No	Nonlinear	No	-	-
Ref. [33] in 2011	2-CCIIIs	3-C, 3-R	No	Yes	Yes	No	No	Nonlinear	No	-	-
Ref. [34] in 2012	2-CCII, 1-UVC	3-C, 3-R	Yes	Yes	Yes	No	Yes	Nonlinear	No	-	-
Ref. [35] in 2013	3-DVCC	3-C, 3-R	No	Yes	Yes	No	Yes	Nonlinear	No	-	-
Ref. [36] in 2014	3-OTRA	3-C, 5-R	No	Yes	No	No	No	Nonlinear	No	-	-
Ref. [37] in 2014	3-OTRA	3-C, 5-R	No	Yes	No	No	No	Nonlinear	No	-	-
Ref. [38] in 2005	4-CCCII	3-C	Yes	Yes	No	Yes	Yes	Nonlinear	No	-	-
Ref. [39] in 2010	3-CCCII	3-C	Yes	Yes	Yes	Yes	Yes	Nonlinear	No	-	-
Ref. [40] in 2009	3-CDTA	3-C	Yes	Yes	No	Yes	No	Nonlinear	No	-	-
Ref. [41] in 2010	3-CDTA	3-C	Yes	Yes	Yes	Yes	No	Nonlinear	No	-	-
Ref. [42] in 2011	1-CCCDTA, 1-OTA	3-C	Yes	Yes	Yes	Yes	No	Nonlinear	No	-	-
Ref. [43] in 2009	2-CCCCTA	3-C	Yes	Yes	No	Yes	No	Nonlinear	No	-	-
Ref. [44] in 2011	1-CCCCTA, 1-OTA	3-C	Yes	Yes	No	Yes	No	Nonlinear	No	-	-
Ref. [45] in 2012	2-OTRA	3-C, 4-R	No	No	No	Yes	No	Nonlinear	No	-	-
Ref. [46] in 2013	1-DDCC, 1-VDTA	3-C, 1-R	Yes	Yes	No	Yes	No	Nonlinear	No	-	-
Ref. [47] in 2015	2-DVCCTA	3-C, 2-R	Yes	Yes	Yes	No	Yes	Nonlinear	No	-	~0-4
Ref. [48] in 2011	18-BJT, 18-current source	3-C	Yes	Yes	No	Yes	Yes	Nonlinear	No	-	-
Ref. [49] in 2012	1 MCCCTA	3-C	Yes	Yes	Yes	Yes	Yes	Nonlinear	No	-	-
Ref. [50] in 2014	3-OTRA	3-C, 4-R	No	Yes	No	No	No	Nonlinear	No	-	-
Ref. [51] in 2014	2-DVTA	3-C	Yes	Yes	No	Yes	No	Nonlinear	No	-	-
Ref. [52] in 2014	2-DVTA	3-C	Yes	Yes	Yes	Yes	No	Nonlinear	No	-	-

Tab. 1. Comparison of the proposed third-order QO with those of previous third-order QOs.

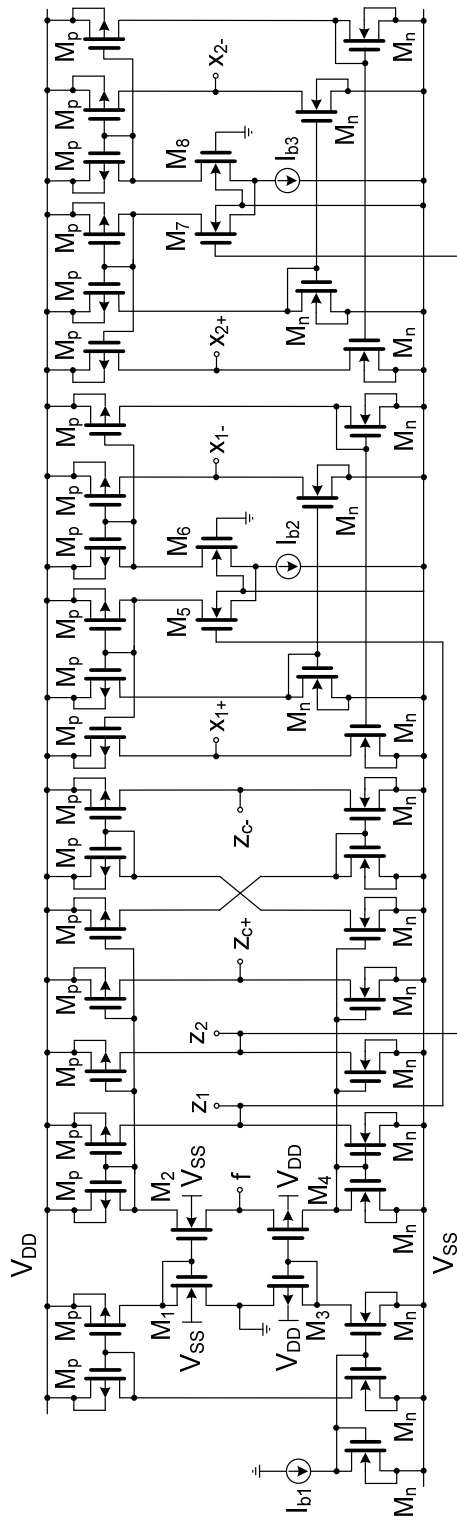


Fig. 2. CMOS implementation for MCCFTA.

The transconductance gains g_{m1} and g_{m2} can be obtained by assuming the transistors M_5 - M_6 and M_7 - M_8 are matched and operated in saturation region, g_{m1} and g_{m2} of MCCFTA can be expressed, respectively, as

$$g_{m1} = \sqrt{\mu C_{ox} \left(\frac{W}{L} \right) I_{b2}}, \quad (3)$$

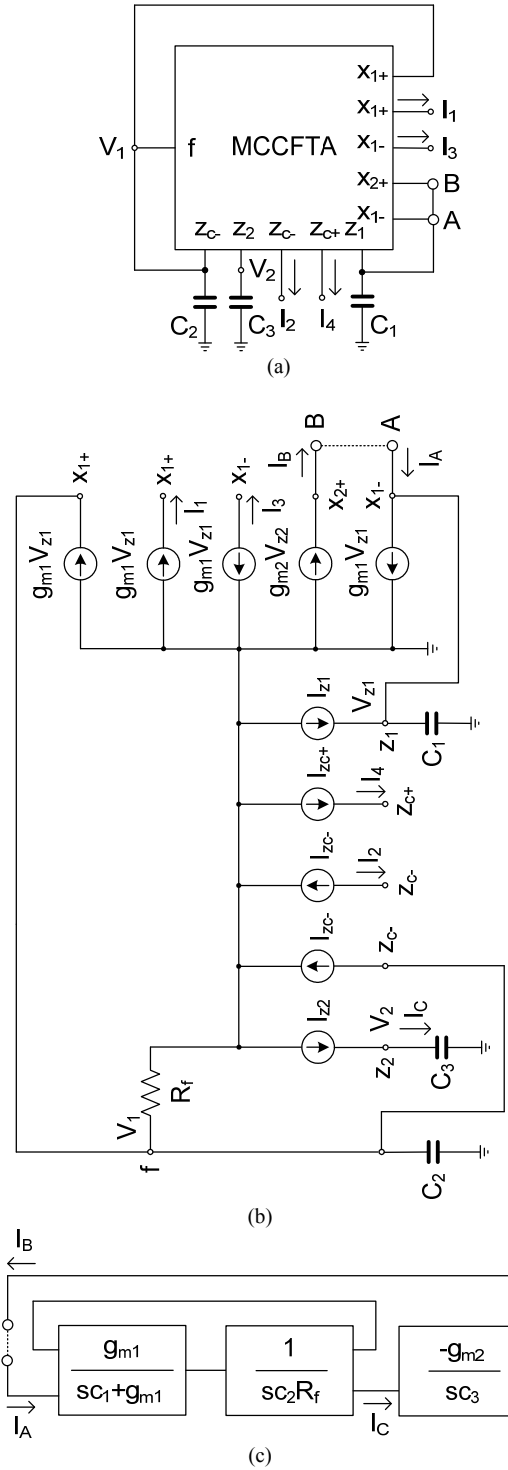


Fig. 3. Proposed third-order QO: (a) circuit symbol, (b) equivalent circuit, (c) block diagram.

$$g_{m2} = \sqrt{\mu C_{ox} \left(\frac{W}{L} \right) I_{b3}}. \quad (4)$$

From (3) and (4), transconductance gains g_{m1} and g_{m2} can be controlled by adjusting the bias currents I_{b2} and I_{b3} , respectively. The z_{c+} - and z_{c-} -terminals (z -copy CFTA) can be obtained by adding current mirrors and cross couple current mirrors in CCFTA as shown in Fig. 2.

The third-order oscillator can be realized from the third-order polynomial equation as

$$N(s) = a_0 s^3 + a_1 s^2 + a_2 s + a_3. \quad (5)$$

Letting $N(s) = 0$ and $s = j\omega$, (5) becomes the third-order oscillator characteristic as

$$0 = -j\omega^3 a_0 - \omega^2 a_1 + j\omega a_2 + a_3. \quad (6)$$

Considering the coefficient of real and imaginary parts, we have:

$$a_3 - \omega^2 a_1 = 0, \quad (7)$$

$$\omega a_2 - \omega^3 a_0 = 0. \quad (8)$$

From (7) and (8), the CO and FO can be given respectively [30] by

$$a_0 a_3 = a_1 a_2, \quad (9)$$

$$\omega = \sqrt{\frac{a_3}{a_1}} = \sqrt{\frac{a_2}{a_0}}. \quad (10)$$

The proposed mixed-mode third-order QO is shown in Fig. 3. It is continuously developed from the circuit proposed in [56] by expanding text, simulation results and adding experimental results. The circuit is consisted of only one MCCFTA and three grounded capacitors. The use of grounded capacitors makes the proposed circuit ideal for IC implementation [57]. The characteristic equation of proposed circuit can be expressed using Fig. 3(b). To easy following, the proposed QO can be shown as block diagram in Fig. 3(c). Considering currents I_A and I_C , current transfer function between I_C and I_A can be given by

$$\frac{I_C}{I_A} = \frac{g_{m1}}{s^2 C_1 C_2 R_f + s C_2 g_{m1} R_f + g_{m1}}. \quad (11)$$

In addition, considering currents I_B and I_A , current transfer function can be given by

$$\frac{I_B}{I_A} = \frac{-g_{m1} g_{m2}}{s^3 C_1 C_2 C_3 R_f + s^2 C_2 C_3 g_{m1} R_f + s C_3 g_{m1}}. \quad (12)$$

Letting $I_B/I_A = 1$ (nodes A and B are closed), the characteristic equation of proposed QO in Fig. 3 can be expressed as

$$s^3 + \frac{g_{m1}}{C_1} s^2 + s \frac{g_{m1}}{C_1 C_2 R_f} + \frac{g_{m1} g_{m2}}{C_1 C_2 C_3 R_f} = 0. \quad (13)$$

Compared with (5), the coefficients can be expressed as:

$$\begin{aligned} a_0 &= 1, \\ a_1 &= \frac{g_{m1}}{C_1}, \\ a_2 &= \frac{g_{m1}}{C_1 C_2 R_f}, \\ a_3 &= \frac{g_{m1} g_{m2}}{C_1 C_2 C_3 R_f}. \end{aligned}$$

According to (9) and (10), the CO and FO are obtained, respectively, by

$$g_{m2} = \frac{g_{m1} C_3}{C_1}, \quad (14)$$

$$\omega_o = \sqrt{\frac{g_{m1}}{C_1 C_2 R_f}}. \quad (15)$$

It is evident from (14) and (15) that the CO can be controlled using g_{m2} by keeping $C_1 = C_3$ and g_{m1} constant and the FO can be controlled by R_f by keeping $C_1 = C_2$ ($C_1 = C_2 = C_3$) and g_{m1} constant. Therefore, the CO and FO can be orthogonally controlled. From Fig. 3, the relationship between I_2 and I_1 can be expressed by

$$\frac{I_2}{I_1} = \frac{1}{s C_2 R_f} \quad (16)$$

while the relationship between voltages V_1 and V_2 can be expressed as

$$\frac{V_2}{V_1} = \frac{1}{s C_3 g_{m2}} \quad (17)$$

where the phase difference is $\phi = \pi/2$. This guarantees that the proposed QO provides the quadrature output currents I_1 and I_2 and quadrature output voltages V_1 and V_2 . Also the uses of multiple-output MCCFTA that provides inversion of the output currents, thus it leads to $I_3 = -I_1$ and $I_4 = -I_2$. Moreover, all current output terminals are at high impedance of MCCFTA, thus ensuring insensitive current outputs that require no additional current followers to be sensed. However, if quadrature output voltages V_1 and V_2 are used, loads cannot be connected directly, the buffer circuit is needed. This problem can be solved easily using voltage follower circuit.

3. Non-Ideal Analysis

To consider the non-ideal effect of a MCCFTA by taking the non-idealities of the MCCFTA into account, the relationship of the terminal voltages and currents can be rewritten as

$$\begin{pmatrix} I_{z1} \\ I_{z2} \\ I_{zc} \\ V_f \\ I_{x1} \\ I_{x2} \end{pmatrix} = \begin{pmatrix} \alpha_1 & 0 & 0 & 0 & 0 & 0 \\ \alpha_2 & 0 & 0 & 0 & 0 & 0 \\ \alpha_c & 0 & 0 & 0 & 0 & 0 \\ R_f & 0 & 0 & 0 & 0 & 0 \\ 0 & \pm g_{m1} & 0 & 0 & 0 & 0 \\ 0 & 0 & \pm g_{m2} & 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} I_f \\ V_{z1} \\ V_{z2} \\ V_{zc} \\ V_{x1} \\ V_{x2} \end{pmatrix} \quad (18)$$

where $\alpha_1 = 1 - \varepsilon_{i1}$ and $\varepsilon_{i1} (\varepsilon_{i1} \ll 1)$ is the current tracking error from f-terminal to z_1 -terminal, $\alpha_2 = 1 - \varepsilon_{i2}$ and $\varepsilon_{i2} (\varepsilon_{i2} \ll 1)$ is the current tracking error from f-terminal to z_2 -terminal and $\alpha_c = 1 - \varepsilon_{ic}$ and $\varepsilon_{ic} (\varepsilon_{ic} \ll 1)$ is the current tracking error from f-terminal to z_c -terminal, R_f is the parasitic resistances at f-terminal. At high-frequency operating,

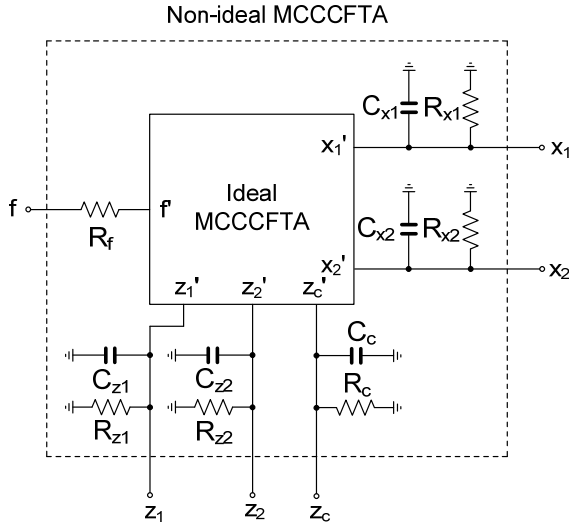


Fig. 4. Non-ideal of MCCFTA at high operating frequency.

non-ideal model of MCCFTA can be shown in Fig. 4. R_{z1} , R_{z2} and R_{zc} are respectively the high parasitic resistances at z_1 -, z_2 - and z_c -terminals, R_{x1} and R_{x2} are respectively the high parasitic resistances at x_1 - and x_2 -terminals, C_{z1} , C_{z2} and C_{zc} are respectively the low parasitic capacitances at z_1 -, z_2 - and z_c -terminals and C_{x1} and C_{x2} are respectively the low parasitic capacitances at x_1 - and x_2 -terminals.

Taking into account the non-ideal MCCFTA characteristics, current transfer function between I_B and I_A can be rewritten as

$$\frac{I_B}{I_A} = \frac{-Z_1 Z_2 Z_3 g_{m1} g_{m2} \alpha_1 \alpha_2 \alpha_c}{(Z_1 Z_2 g_{m1} + Z_1 R_f g_{m1} + Z_2 + R_f + Z_1 Z_2 g_{m1} \alpha_1 \alpha_c - Z_1 Z_2 g_{m1} \alpha_1 - Z_2 \alpha_1)} \quad (19)$$

where

$$\begin{aligned} Z_1 &= \frac{R_{p1}}{sC_1' R_{p1} + 1}, \\ Z_2 &= \frac{R_{p2}}{sC_2' R_{p2} + 1}, \\ Z_3 &= \frac{R_{p3}}{sC_3' R_{p3} + 1}, \\ C_1' &= C_1 \parallel C_{x1-} \parallel C_{x2+} \parallel C_z, \\ C_2' &= C_2 \parallel C_{x1+} \parallel C_{x2-}, \\ C_3' &= C_3 \parallel C_{z2}, \\ R_{p1} &= R_{x1-} \parallel R_{x2+} \parallel R_z, \\ R_{p2} &= R_{x1+} \parallel R_{x2-}, \\ R_{p3} &= R_{z2}. \end{aligned}$$

The modified characteristic equation of Fig. 3 can be rewritten as

$$\begin{aligned} s^3 + s^2 \frac{g_{m1}}{C_1'} &\left(1 + \frac{1}{g_{m1} R_f} + \frac{1}{g_{m1} R_{p2}} + \frac{1}{g_{m1} R_{p1}} \right) \\ &+ \frac{C_1'}{C_3' g_{m1} R_{p3}} - \frac{\alpha_1}{g_{m1} R_f} \\ &+ s \frac{g_{m1}}{C_1' C_2' R_f} + \frac{C_1'}{C_3' g_{m1} R_{p3}} + \frac{R_f}{g_{m1} R_{p1} R_{p2}} + \frac{C_1' R_f}{C_3' g_{m1} R_{p2} R_{p3}} \\ &+ \frac{C_2' R_f}{C_3' g_{m1} R_{p1} R_{p3}} - \frac{\alpha_1}{g_{m1} R_{p1}} - \frac{\alpha_1}{g_{m1} R_{p1}} - \alpha_1 \\ &+ \frac{g_{m1} g_{m2} \alpha_1 \alpha_2 \alpha_c}{C_1' C_2' C_3' R_f} \left(1 + \frac{1}{g_{m1} g_{m2} R_{p1} R_{p3} \alpha_1 \alpha_2 \alpha_c} + \frac{1}{g_{m2} R_{p3} \alpha_2} \right. \\ &+ \frac{1}{g_{m2} R_{p3} \alpha_1 \alpha_2 \alpha_c} + \frac{R_f}{g_{m2} R_{p2} R_{p3} \alpha_1 \alpha_2 \alpha_c} \\ &+ \frac{R_f}{g_{m1} g_{m2} R_{p1} R_{p2} R_{p3} \alpha_1 \alpha_2 \alpha_c} \\ &\left. - \frac{1}{g_{m1} g_{m2} R_{p1} R_{p3} \alpha_2 \alpha_c} - \frac{1}{g_{m2} R_{p3} \alpha_2 \alpha_c} \right). \quad (20) \end{aligned}$$

Letting, parasitic parameters R_z , R_c , R_x are very high resistance values and C_z , C_c , C_c are very low capacitance value, (20) can be approximated as

$$s^3 + s^2 \frac{g_{m1}}{C_1'} + s \frac{g_{m1}}{C_1' C_2' R_f} + \frac{g_{m1} g_{m2} \alpha_1 \alpha_2 \alpha_c}{C_1' C_2' C_3' R_f} = 0. \quad (21)$$

In this case, the CO and FO are modified, respectively, as:

$$g_{m2} = \frac{g_{m1} C_3'}{C_1'} \alpha_1 \alpha_2 \alpha_c, \quad (22)$$

$$\omega_o = \sqrt{\frac{g_{m1} g_{m2} \alpha_1 \alpha_2 \alpha_c}{C_1' C_2' C_3' R_f}}. \quad (23)$$

The various passive sensitivities of ω_o of the proposed QO can be obtained as

$$\left. \begin{aligned} S_{g_{m1}}^{\omega_o} &= S_{g_{m2}}^{\omega_o} = S_{\alpha_1}^{\omega_o} = S_{\alpha_2}^{\omega_o} = S_{\alpha_3}^{\omega_o} = 0.5 \\ S_{R_f}^{\omega_o} &= S_{C_1'}^{\omega_o} = S_{C_2'}^{\omega_o} = S_{C_3'}^{\omega_o} = -0.5 \end{aligned} \right\} \quad (24)$$

Thus, the proposed QO has low active and passive sensitivities.

4. Simulation Results

To verify the theoretical prediction of the proposed third-order QO, the circuit in Fig. 3 was simulated using PSPICE simulators and the MCCFTA in Fig. 2 was used. The model parameters for nMOS and pMOS transistors are taken from 0.25 μm CMOS process from TSMC. The power supply was given as ± 1 V. From Fig. 2, I_{b1} was

initially designed for 100 μA while I_{b2} and I_{b3} were initially designed for 150 μA . Transistors around trans-linear loop (M_1 to M_4) were designed to obtain $g_{m(\text{nMOS})} = g_{m(\text{pMOS})}$ and differential transistor pairs (M_5 - M_6 and M_7 - M_8) were designed to obtain $g_{m(\text{total})} = (g_{m(M5)} + g_{m(M6)})/2$ or $g_{m(\text{total})} = (g_{m(M7)} + g_{m(M8)})/2$. For 0.25 μm CMOS process used in this case, $(\mu_o C_{ox})_{\text{nMOS}} = 242.2 \mu\text{A/V}^2$ and $(\mu_o C_{ox})_{\text{pMOS}} = 51.8 \mu\text{A/V}^2$, $V_{th0(\text{nMOS})} = 0.37 \text{ V}$ and $V_{th0(\text{pMOS})} = -0.49 \text{ V}$ were given. Therefore, the aspect ratios of MOS transistors for MCCFTA in Fig. 3 were $W/L = 5 \mu\text{m}/0.5 \mu\text{m}$ for M_1 - M_2 , $W/L = 15 \mu\text{m}/0.5 \mu\text{m}$ for M_3 - M_4 , $W/L = 5 \mu\text{m}/0.5 \mu\text{m}$ for M_5 - M_8 , $W/L = 5 \mu\text{m}/0.5 \mu\text{m}$ for all M_n transistors and $W/L = 15 \mu\text{m}/0.5 \mu\text{m}$ for all M_p transistors. PSPICE simulations have verified that when I_{b1} was varied from 1 to 100 μA , achieved R_f was in the range of 12 to 2.19 $\text{k}\Omega$ as shown in Fig. 5 whereas I_{b2} and I_{b3} were varied from 5 to 150 μA , achieved g_{m1} and g_{m2} were in the range of 107 to 472 $\mu\text{A/V}$ as shown in Fig. 6. Both simulated R_f - and g_m - value were compared with (2) and (3), respectively. Simulated performances of MCCFTA were summarized in Tab. 2.

The voltage outputs and current outputs were shown respectively in Figs. 7 and 8, when Fig. 3 was designed with the following values: $C_1 = C_2 = C_3 = 20 \text{ pF}$, $I_{b1} = 20 \mu\text{A}$, $I_{b2} = 80 \mu\text{A}$ and $I_{b3} = 75 \mu\text{A}$. The I_{b3} -value was varied for g_{m2} so as to satisfy the CO in (14). In this case, ($R_f \approx 1/g_{m1}$), each current output and each voltage output were almost equal of magnitude. The magnitudes of the quadrature signals will not be equaled in case of R_f was varied ($R_f \neq 1/g_{m1}$) in (15). For applications requiring equal magnitude quadrature outputs, additional amplifying circuits were needed. From Fig. 7, the simulated THD of V_1 and V_2 were about 0.26 % and 0.17 %, respectively, the FO was 2.7 MHz. From Fig. 8, the simulated the FO was found to be 2.7 MHz, the results of the I_1 , I_2 , I_3 and I_4 total harmonic distortion (THD) analysis were obtained as 0.56 %, 0.37 %, 0.85 % and 0.38 %, respectively, and the power consumption was about 1.43 mW. This low THD was obtained for the output signals V_1 , V_2 , I_1 , I_2 , I_3 and I_4 of 204 mV_{P-P}, 198 mV_{P-P}, 70 μA_{P-P} , 70 μA_{P-P} , 70 μA_{P-P} and

Parameters	Value
Technology	0.25 μm CMOS
Power supply	$\pm 1 \text{ V}$
R_f ($I_b = 1$ -100 μA)	12 to 2.19 $\text{k}\Omega$
g_m ($I_b = 5$ -150 μA)	107 to 472 $\mu\text{A/V}$
Input and output range ($I_b = 5$ -150 μA)	± 40 to $\pm 250 \text{ mV}$
Bandwidth (-3dB) @ $I_{b1} = 50 \mu\text{A}$, $I_{b2} = I_{b3} = 100 \mu\text{A}$	42.84 MHz 39.9 MHz
Current follower (I_z/I_f)	
Current follower (I_x/I_f) ($R_z = 2.23 \text{ k}\Omega$)	1.034
Current gain @ $I_{b1} = 50 \mu\text{A}$, $I_{b2} = I_{b3} = 100 \mu\text{A}$ - I_z/I_f	
Parasitic parameters @ $I_{b1} = 50 \mu\text{A}$, $I_{b2} = I_{b3} = 100 \mu\text{A}$	116 $\text{k}\Omega$, 25 fF 96 $\text{k}\Omega$, 25 fF
- R_z , C_z - R_x , C_x	
Power consumption @ $I_{b1} = 50 \mu\text{A}$, $I_{b2} = I_{b3} = 100 \mu\text{A}$	2.12 mW

Tab. 2. Simulated specifications of MCCFTA.

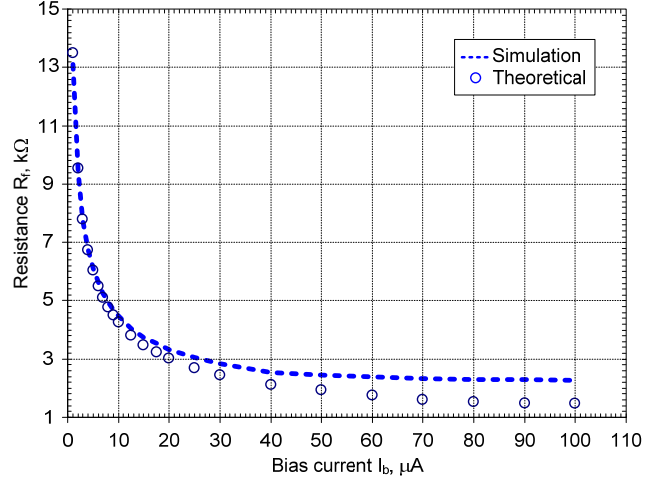


Fig. 5. Simulated resistance R_f when bias current is varied.

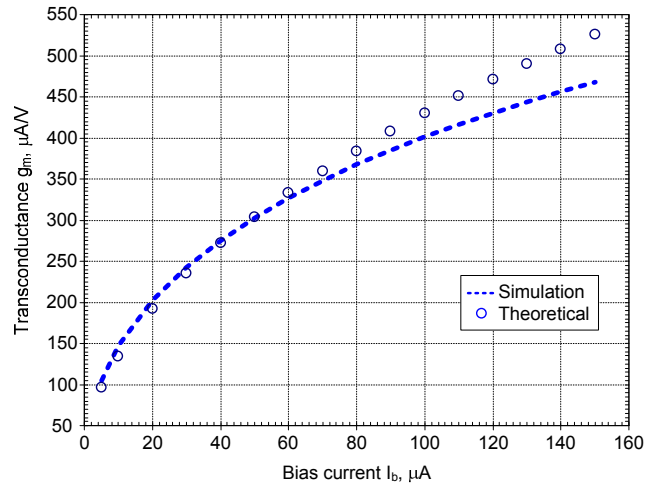


Fig. 6. Simulated transconductance g_m when bias current is varied.

66 μA_{P-P} , respectively. THD will be increased if output signal level was increased which can be obtained by increased CO. However, increasing CO may be increased nonlinear behavior of the system which is results in high THD output signal that should be avoided.

The electronic tuning of the oscillator was shown in Fig. 9. The result gives a variation of the FO from 1.1 to 3.25 MHz with I_{b1} in the range of 1 to 50 μA that was confirmed by (15). This FO was achieved without adjusting the CO. It shows that the simulated FO was consistent with the theoretical values. At $I_{b1} = 50 \mu\text{A}$, the error of FO between the simulated and theoretical values was 4.9 %. The bias current I_{b1} can be continuously varied high up to 100 μA , but the g_{m2} must be adjusted via I_{b3} to satisfy the CO. However, it was also found that there was a deviation between the theoretical and simulated values in large bias current value region over the value of 50 μA . From Fig. 9, output signal levels of V_1 , V_2 , I_1 , I_2 , I_3 , and I_4 versus f_o have been obtained as shown in Fig. 10. It should be noted that the output signal level is not equal and constant when I_{b1} was varied. This problem can be solved by using automatic gain control (AGC) circuit. There are oscillators with AGC available in literature [58–61]. These techniques use AGC

circuit including into the loop of the system for controlling the amplitude of signal. Usually, AGC is suitable for applying to linear control of CO and FO oscillator, if nonlinear control of CO and FO such as this work (adjusting CO and FO in square-root domain), current squaring circuits [62], [63] are required to compensate the square-root form to obtain linear current control of CO and FO.

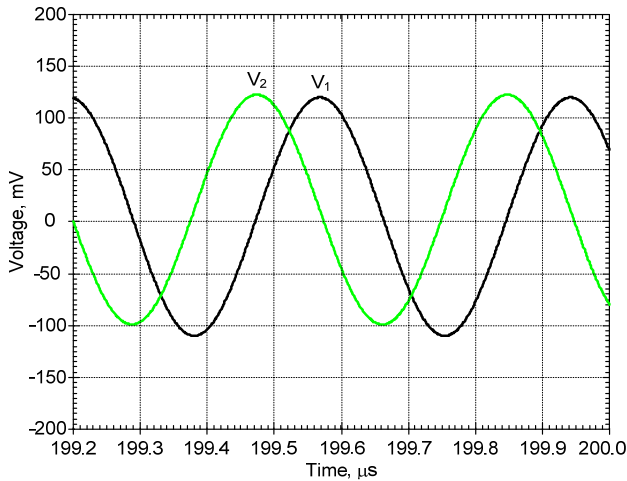


Fig. 7. Simulated two voltage output waveforms.

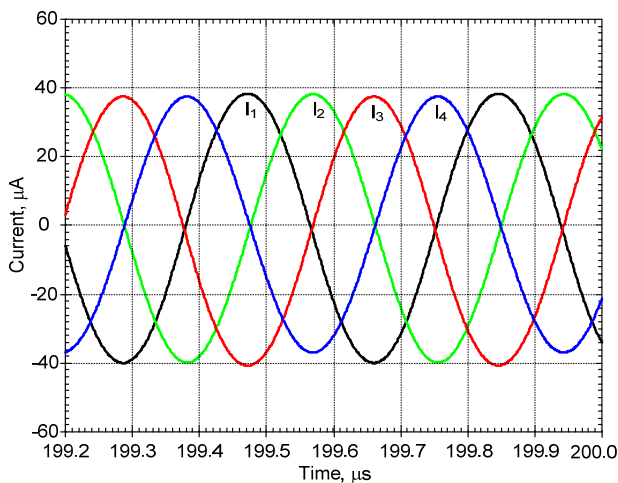


Fig. 8. Simulated four current output waveforms.

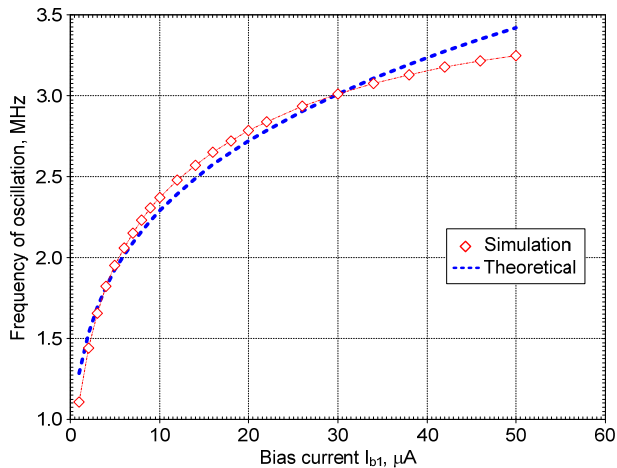


Fig. 9. Electronic frequency tuning with the bias current I_{b1} .

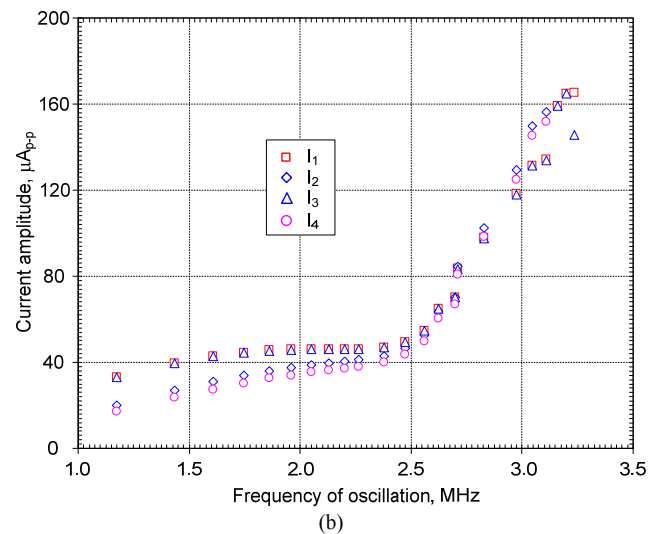
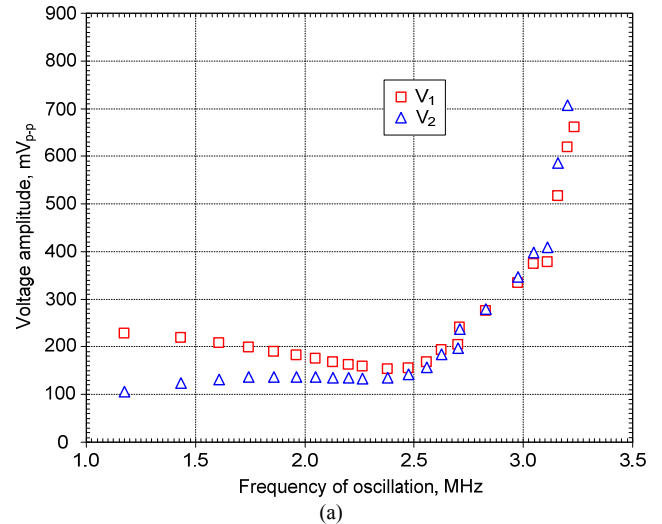


Fig. 10. Output level versus f_o : (a) V_1 and V_2 , (b) I_1 , I_2 , I_3 and I_4 .

Total harmonic distortion (THD) and phase error were shown in Figs 11 and 12, respectively. From Fig. 11, THD was increased when I_{b1} (R_f) was varied far from satisfying condition.

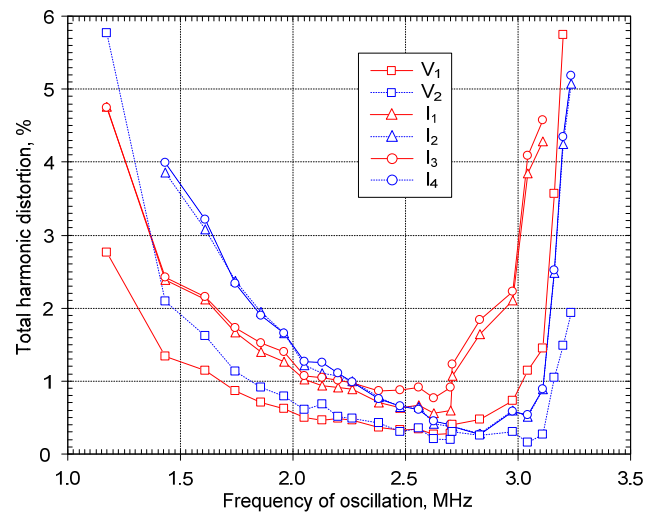


Fig. 11. Simulated THD versus f_o .

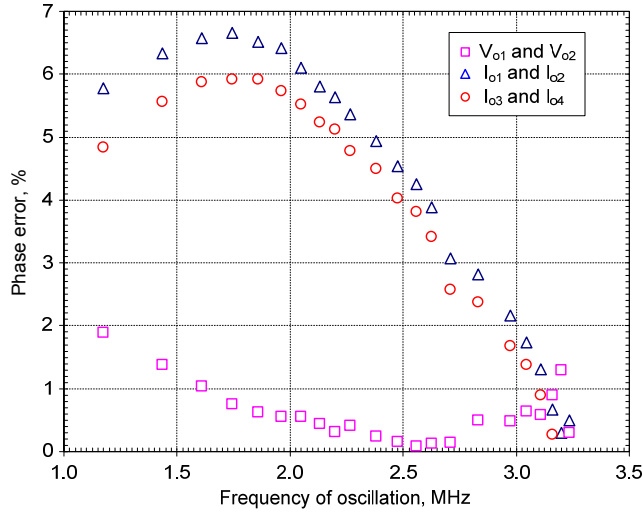


Fig. 12. Simulated phase error versus f_o .

5. Experimental Results

The proposed oscillator was experimentally tested wherein the functionality of the MCCFTA was implemented using CFOA AD844 and OTA LM13600 as shown in Fig. 13. The supply voltage ± 10 V was used. The non-instrument of oscilloscope as measurable currents is available to the authors. However, for sake of experimental results, the current outputs, I_1 , I_2 , I_3 and I_4 , can be obtained by connecting the external resistors. In this experiment, current outputs I_1 , I_2 , I_3 and I_4 were connected to the resistors $50\text{ k}\Omega$ and voltage across these resistors will be measured. To obtain the experimental results, Tektronix MSO

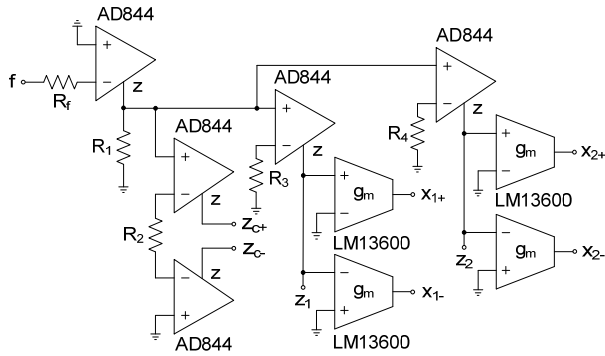


Fig. 13. Possible realization of the MCCFTA using commercial active devices (AD844 and LM13600).

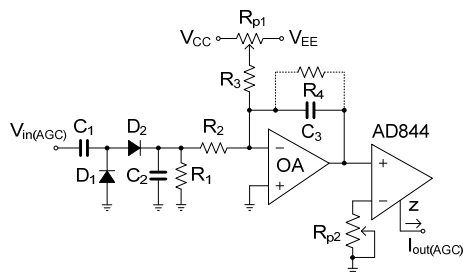


Fig. 14. Amplitude-automatic gain control circuit [61].

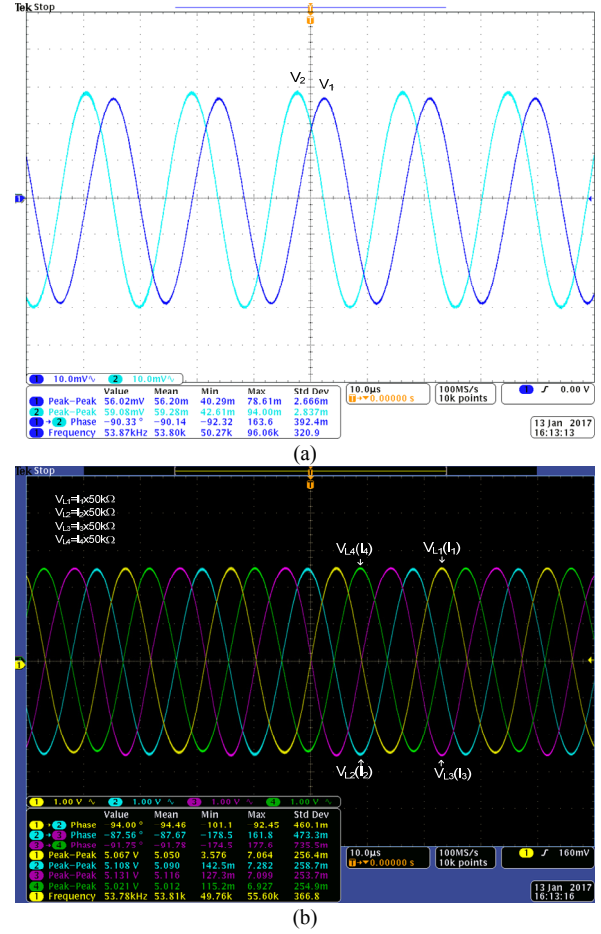


Fig. 15. Measured waveforms for: (a) voltage outputs, (b) current outputs.

4034 oscilloscope and Keysight N9030A spectrum analyzer were used. The capacitors were designed with $C_1 = C_2 = C_3 = 5\text{ nF}$. The resistors R_1 , R_2 , R_3 and R_4 in Fig. 13 were given as $1\text{ k}\Omega$. The variable resistor was used to obtain the effect of variable g_m and R_f .

To control the output amplitude of signals to be constant, the amplitude-automatic gain control (AGC) circuit in Fig. 14 was used. This circuit was adopted from [61]. From our pretest on the proposed circuit, it was found that V_1 (I_1 and I_3) was dependent on tuning of FO, thus V_1 will be used for the input of AGC circuit ($V_{in(AGC)}$). CFOA AD844 and R_{p2} were used to work as voltage-to-current converter ($V-I$ converter). The current output $I_{out(AGC)}$ of AGC circuit will be supplied additionally to bias current I_{b3} for compensating CO. The active device and passive-value used in Fig. 14 were tabulated in Tab. 3.

The quadrature sinusoidal voltage and current output waveforms with resistor $R_f = 500\text{ }\Omega$, bias currents $I_{b2} = 50\text{ }\mu\text{A}$ ($g_{m2} = 1\text{ mA/V}$) and $I_{b3} = 61.8\text{ }\mu\text{A}$ were shown in Fig. 15, (a) and (b), respectively. It should be noted that the quadrature sinusoidal signal was almost equal of amplitude. The FO was found as 53 kHz . From Fig. 15, the THDs for V_1 , V_2 , I_1 , I_2 , I_3 and I_4 were about 1.2, 0.9, 1.2, 1.1, 1.3 and 1.1 %, respectively. The quadrature relationship was further verified through the X-Y plots of the two output forms in Fig. 15 as shown in Fig. 16, (a) and (b), respectively.

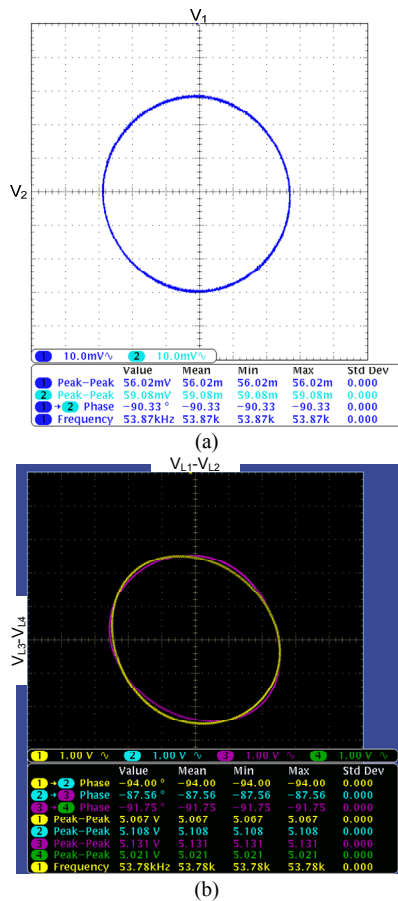


Fig. 16. X-Y plots for: (a) voltage output, (b) current output.

Device	Value
OA	OPA2650
D ₁ , D ₂	2xBAT42
C ₁ , C ₂	1 μ F
C ₃	10 nF
R ₁ , R ₄	1 M Ω
R ₂ , R ₃	100 k Ω
R ₅	200 Ω
R _{p1} and R _{p2}	100 k Ω (variable resistor)

Tab. 3. Active and passive components used in Fig. 14.

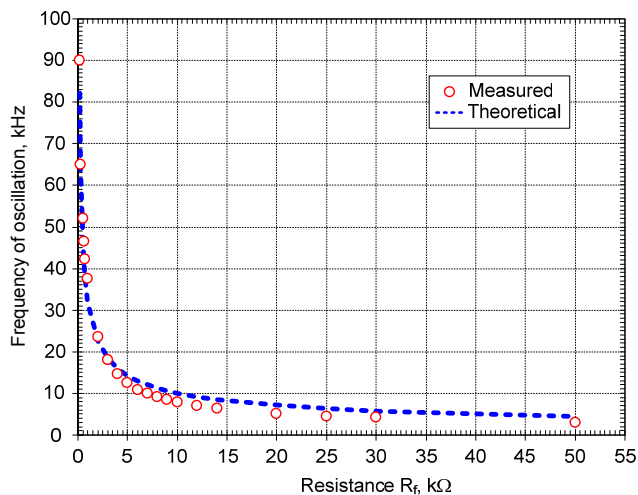
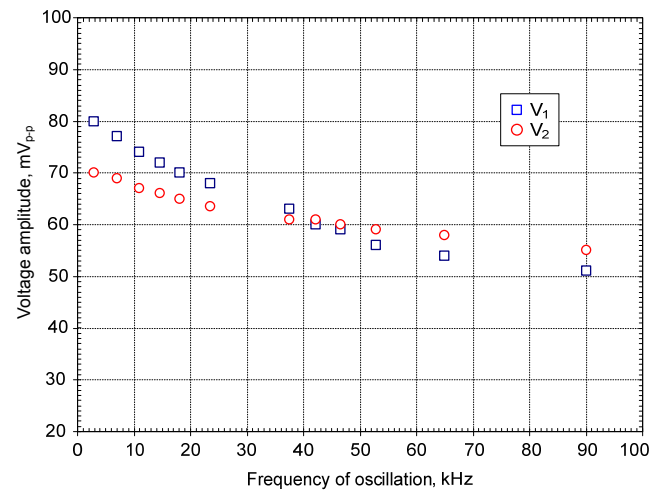
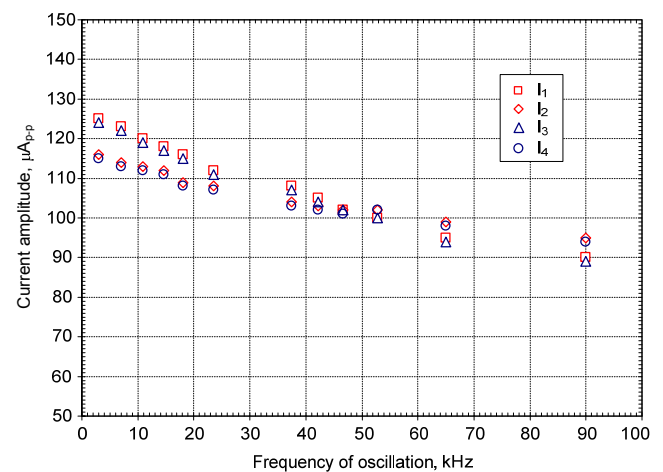
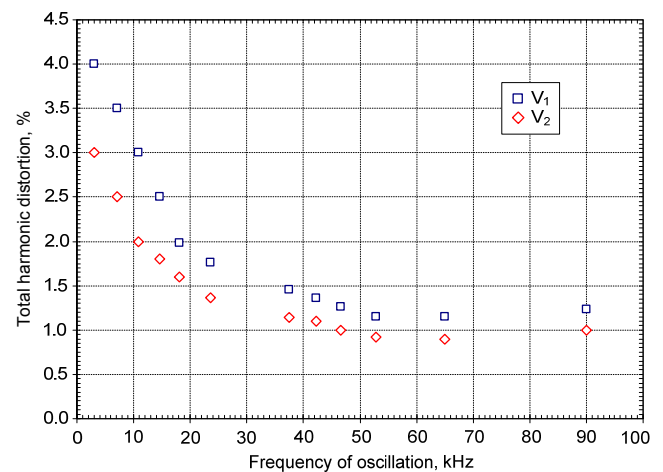

 Fig. 17. FO tuning with R_f .

 Fig. 18. Output level of V_1 and V_2 versus f_0 .

 Fig. 19. Output level of I_1 , I_2 , I_3 and I_4 versus f_0 .

 Fig. 20. Measured THD of voltage outputs versus f_0 .

Figure 17 shows the experimental results of the FO by changing the value of the resistor R_f with $C_1 = C_2 = C_3 = 5$ nF, $I_{b1} = 50$ μ A and $I_{b3} = 61.8$ μ A. The tuning of the oscillator gives a variation of the FO from 2.9 to 90 kHz when R_f -value was decreased from 50 to 0.15 k Ω . In this case, FO was obtained by adjusting CO via AGC. The plots for theoretical value were also included for comparison.

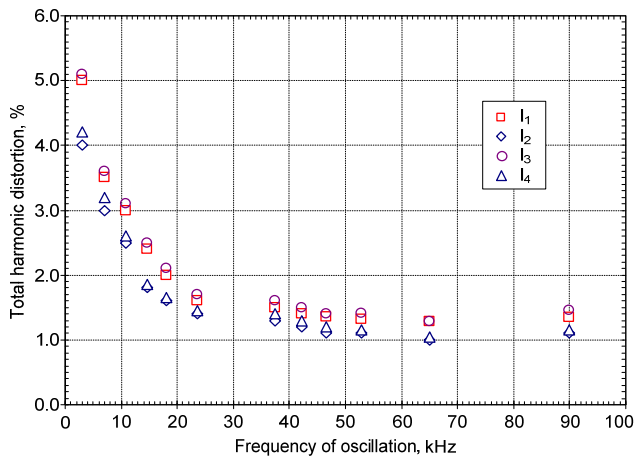


Fig. 21. Measured THD of current outputs versus f_o .

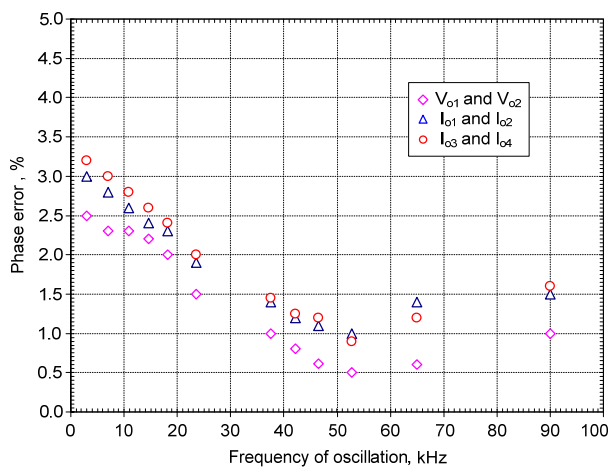


Fig. 22. Measured phase error versus f_o .

From Fig. 17, output signal level of V_1 and V_2 versus f_o and output signal level of I_1 , I_2 , I_3 and I_4 versus f_o were shown in Fig. 18 and 19, respectively. It should be noted that the amplitude was not fluctuated; thanks to AGC for obtaining this result. THDs for voltage outputs and current outputs versus f_o were measured and shown in Figs. 20 and 21. Finally, phase errors versus f_o were also measured and shown in Fig. 22.

Compared between the simulation and experimental results, setting of FO-value and working capacitor-value were different. Actually, this work focuses on the QO which should be implemented as IC form. Unfortunately, the obstruction for fabricating was the cost. Although, higher FO for experimental results was possible, but our test was investigated the circuit which builds on perfboard. Using printed circuit board and high quality active devices the operation band of the circuit [59–61] will be increased. Therefore, this experimental result was used only to confirm the workability of the proposed structure.

6. Conclusions

In this paper, a mixed-mode third-order QO based on new MCCFTA has been presented. The proposed circuit

uses only one MCCFTA and three grounded capacitors. The use of grounded capacitors is ideally interesting from an integration point of view. The proposed structure provides four high output impedance current sources with 90° phase difference, thus these output terminals can be directly connected to the loads without additional follower circuits. In addition, voltage with 90° phase difference can be obtained without changing any topology. Also the CO and FO can be controlled orthogonally and electronically by adjusting the bias currents of MCCFTA. The active and passive sensitivities are no more than unity of magnitude. Simulation and experimental results, which confirm the theoretical analysis, are included.

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